TITLE

METHOD AND APPARATUS FOR I/Q MISMATCH CALIBRATION OF TRANSMITTER

BACKGROUND OF THE INVENTION

The present invention relates to compensation of physical
5   layer impairments on transmitter systems and particularly to a
method and apparatus for I/Q mismatch calibration in transmitter
systems.

Description of the Related Art
10   OFDM is a multi-channel modulation system employing
Frequency Division Multiplexing (FDM) of orthogonal
sub-carriers, each modulating a low bit-rate digital stream. The
simplest way to describe an orthogonal frequency-division
multiplexing (OFDM) signal is as a set of closely spaced
15   frequency-division multiplexed carriers. While this is a good
starting point for those unfamiliar with the technology, it falls
short as a model for analyzing the effects of signal impairment.

The reason it falls short is that the carriers are more than
closely spaced; they are heavily overlapped. In a perfect OFDM
20   signal, the orthogonality property prevents interference between
overlapping carriers. This is different from the FDM systems.
In FDM systems, any overlap in the spectrums of adjacent signals
will result in interference. In OFDM systems, the carriers will
interfere with each other only if there is a loss of
25   orthogonality. So long as orthogonality can be maintained, the
carriers can be heavily overlapped, allowing increased spectral
efficiency.

1

Table 1 lists a variety of common analog signal impairments and their effects on both OFDM signals and the more familiar single-carrier modulations such as quadrature phase-shift keying (QPSK) or 64-QAM (quadrature amplitude modulation). Most of

5 these impairments can occur in either the transmitter or the receiver.

Impairment	OFDM	QPSK
I/Q gain balance	State spreading (uniform/carrier)	Distortion of constellation
I/Q quadrature skew	State spreading (uniform/carrier)	Distortion of constellation
I/Q channel mismatch	State spreading (non-uniform/carrier)	State spreading
Uncompensated frequency error	State spreading	Spinning constellation
Phase noise	State spreading (uniform/carrier)	Constellation phase arcing
Nonlinear distortion	State spreading	State spreading
Linear distortion	Usually no effect (equalized)	State spreading if not equalized
Carrier leakage	Offset constellation for center carrier only (if used)	Offset constellation
Frequency error	State spreading	Constellation phase arcing
Amplifier droop	Radial constellation distortion	Radial constellation distortion
Spurious	State spreading or shifting of affected sub-carrier	State spreading, generally circular

For cost reasons, analog in-phase and quadrature (I/Q)

10 modulators and demodulators are often used in transceivers especially for wide bandwidth signals. Being analog, these I/Q modulators and demodulators usually have imperfections that result in an imperfect match between the two baseband analog signals, I and Q, which represent the complex carrier. For

15 example, gain mismatch might cause the I signal to be slightly different from the Q. In a single-carrier modulation system, this

2

results in a visible distortion in the constellation— the square constellation of a 64-QAM signal would become rectangular.

To better understand how gain imbalance will affect an OFDM signal, look at the equations describing each individual sub-carrier. In the following analysis, it's important to keep in mind that, while an individual sub-carrier is analyzed, the I/Q gain imbalance error is on the signal that is the composite of all sub-carriers.

In the equation (1), $C_{k,m}$ is a complex number representing the location of the symbol within the constellation for the kth sub-carrier at the mth symbol time. For example, if sub-carrier k is binary-phase-shift-keying (BPSK) modulated, then $C_{k,m}$ might take on values of $\pm 1 + j0$. The complex exponential portion of equation (1) represents the kth sub-carrier, which is amplitude- and phase-modulated by the symbol $C_{k,m}$. Therefore:

$$C_{k,m}(e^{j2\pi k\Delta ft}) \dotfill (1)$$

where Δf is the frequency space between the sub-carrier. Using Euler's relation, the equation (1) can be rewritten as:

$$C_{k,m}(\cos(2\pi \cdot k\Delta ft) + j\sin(2\pi \cdot k\Delta ft)) \dotfill (2)$$

Now add a term "β" to represent gain imbalance. For a perfect signal, set $\beta = 0$. As shown, the gain imbalance term will also produce a gain change. This was done to simplify the analysis. Therefore:

$$C_{k,m}((1 + \beta)\cos(2\pi \cdot k\Delta ft) + j\sin(2\pi \cdot k\Delta ft) \dotfill (3)$$

The equation can be rearranged and this can be rewritten as the sum of a perfect signal and an error signal:

$$C_{k,m}(\cos(2\pi \cdot k\Delta ft) + j\sin(2\pi \cdot k\Delta ft)) + C_{k,m}\beta\cos(2\pi \cdot k\Delta ft) \dotfill (4)$$

Finally, converting back into complex exponential notation, we get:

$$C_{k,m}e^{j2\pi k\Delta ft} + (C_{k,m}\frac{\beta}{2})\cdot(e^{j2\pi k\Delta ft} + e^{-j2\pi k\Delta ft}) \dotfill (5)$$

In words, the equation (5) shows that a gain imbalance produces two error terms. The first error term is at the frequency of the kth sub-carrier. The second error term is at the frequency of the $-k$th sub-carrier. The phase and magnitude of the error terms are proportional to the symbol being transmitted on the kth sub-carrier. Another way of saying this is that I/Q gain imbalance will result in each sub-carrier being interfered with by its frequency mirror-image sub-carrier. Persons skilled in the art will instantly recognize this as imperfect sideband cancellation.

The equation (5) has several implications. First, it is generally true that for sub-carriers used to carry data (as opposed to pilots), the symbol being transmitted at any given time on the kth sub-carrier is uncorrelated to the symbol on the $-k$th sub-carrier.

For a given sub-carrier, the lack of correlation from the mirror-image sub-carrier implies a certain randomness to the error. This results in a spreading of the sub-carrier's constellation states in a noise-like fashion. This is especially true for higher-order modulations such as 64-QAM. For lower-order modulations, such as BPSK, the error term from the mirror-image carrier has fewer states.

This can result in constellations where the BPSK pilot carriers of an 802.11a signal exhibit spreading that does not appear noise-like. Also, as the BPSK pilots do not have an imaginary component; the error terms associated with the pilot sub-carriers are real— so the spreading is only along the real (I) axis. Note that the phase relationships between the pilot

carriers in an 802.11a system are highly correlated, so the errors introduced by quadrature errors are not random.

Quadrature skew produces error terms similar to those produced by gain imbalance. Quadrature skew occurs when the two 5 oscillators used in an I/Q modulator or demodulator do not differ by exactly 90°. For a small angular error, it can be shown that the resulting error is orthogonal to the data. This is indicated by the j in front of the error terms in the equation (6). As with gain imbalance, the error generates energy at the kth and $-k$th 10 sub-carriers. Again, the 802.11a BPSK pilots do not have an imaginary component, so the error term, which is now orthogonal, causes spreading along the Q axis. For the QPSK carriers in this example, the error is also orthogonal. However, unlike BPSK, a QPSK constellation doesn't look any different when rotated by 15 90°. (See the equation (6)):

$$C_{k,m}e^{j2\pi k\Delta ft} + j\frac{C_{k,m}\phi}{2}\cdot(e^{j2\pi k\Delta ft} + e^{-j2\pi k\Delta ft}) \dots\dots\dots (6)$$

In both 802.11a and Hiperlan2, a channel estimation sequence is transmitted at the beginning of a burst. This special sequence is used to train the receiver's equalizer. The intended function 20 of the equalizer is to compensate the received signal for multi-path distortion a linear impairment in the signal that is the result of multiple signal paths between the transmitter and the receiver. As the ideal channel estimation sequence is known by the receiver, the receiver can observe the effects of the 25 channel on the transmitted signal and compute a set of equalizer coefficients.

In the transmitter, the channel estimation sequence is created by BPSK modulating all 52 carriers for a portion of the preamble. Not coincidentally, the equalizer consists of 52

complex coefficients— one for each sub-carrier. It should come as no surprise that each sub-carrier in the channel estimation sequence has the greatest influence on the equalizer coefficient computed for that same sub-carrier.

5   The channel estimation sequence, and the receiver algorithms that compute the equalizer coefficients, are not immune from signal impairments. Consider, for example, the effect of I/Q gain imbalance on sub-carriers +26 and −26 of the channel estimation sequence. Recall from equation (5) that each

10   sub-carrier has two error terms: one at the same frequency as the sub-carrier, and one at the mirror image frequency. The I/Q gain imbalance will cause mutual interference between sub-carriers +26 and −26.

     From the IEEE 802.11a standard, the sub-carrier modulation

15   for the channel estimation sequence is defined to be $C_{-26}=1+j0$ and $C_{+26}=1+j0$. Using these values in equation (5), one can easily determine that the two sub-carriers, when combined with the resulting error terms, will suffer an increase in amplitude. The equalizer algorithm will be unable to differentiate the error

20   from the actual channel response, and will interpret this as a channel with too much gain at these two sub-carrier frequencies. The equalizer will incorrectly attempt to compensate by reducing the gain on these sub-carriers for subsequent data symbols.

     The result will be different for other sub-carrier pairs,

25   depending on the BPSK channel estimation symbols assigned to each.

     With QPSK sub-carriers, the equalizer error caused by gain imbalance, or quadrature skew, results in seven groupings in each corner. Each QPSK sub-carrier suffers from QPSK interference

30   from its mirror image. This results in a spreading to four

constellation points in each corner. Each QPSK sub-carrier also suffers from a bi-level gain error introduced by the equalizer. This would produce eight groupings, except that the gain error is such that corners of the groupings overlap at the ideal corner

5   state. Only seven groupings are visible.

I/Q channel mismatch

When the frequency response of the baseband I and Q channel signal paths are different, an I/Q channel mismatch exists. I/Q

10   channel mismatch can be modeled as a sub-carrier-dependent gain imbalance and quadrature skew. I/Q gain imbalance and quadrature skew, as described above, are simply a degenerate form of I/Q channel mismatch in which the mismatch is constant over all sub-carriers. Think of channel mismatch as gain imbalance and

15   quadrature skew as a function of a sub-carrier. It is still generally true that channel mismatch causes interaction between the kth and $-k$th sub-carriers, but that the magnitude of the impairment could differ between the kth and the $(k+n)$th carriers.

In order to eliminate the effects of the previously

20   described impairments on the OFDM systems, various kinds of compensation circuits and methods have been proposed.

U.S. Application Publication No. 20020015450 discloses a method and an arrangement for determining correction parameters used for correcting phase and amplitude imbalance of an I/Q

25   modulator in a transmitter. The transmitter includes an I/Q modulator and a corrector for correcting the phase and amplitude imbalance caused by the I/Q modulator. The arrangement has means for sampling the I/Q-modulated test signal to be transmitted, means for A/D-converting the signal samples taken from the test

30   signal, means for I/Q-demodulating the signal samples digitally

into I- and Q-feedback signals, means for determining the phase
and amplitude imbalance caused by the I/Q modulator on the basis
of the I- and Q-feedback signals, and means for determining the
correction parameters of phase and amplitude on the basis of the
5 determined phase and amplitude imbalance.

U.S. Patent 6,208,698 discloses a quadrature modulator
controlled by an imbalance estimator, as shown in FIG. 1. The
estimator is entirely analog and includes two mixers 120, 121 to
which are applied two carrier signals P1 and P2 derived from a
10 carrier signal P obtained from a local oscillator and from a 90°
phase-shifter 122. A modulating signal X, Y is applied to each
of the mixers 120, 121. The output signals of the mixers 120,
121 are applied to a combiner 123 to constitute a modulated signal
S. The modulated signal S is either an intermediate frequency
15 signal or a microwave frequency signal. The estimator 124 of the
invention includes means 125 for detecting the instantaneous
power Pd of the modulated signal S, means 126, 127 for multiplying
the detected instantaneous power Pd by each of the modulating
signals X, Y, means 128A, 129A for rectifying the signals
20 produced, which rectifier means can be diodes, for example, and
means 130, 131 for integrating the rectified signals, the
integrator means being followed by subtractors 160, 161 supplying
respective signals E1 and E2 proportional to the amplitude of the
modulator imbalance. The subtractors 160 and 161 receive
25 respective analog reference signals REF1 and REF2, which are
usually identical (REF1=REF2), the output signals of the
integrator means 130, 131 being subtracted from REF1 and REF2,
respectively. For example, REF1 and REF2 are equal to twice the
mean amplitude of the modulated signal S. The subtractors 160
30 and 161 generally need to be used only for the purposes of

8

correcting imbalance. The signals E1 and E2 are respectively applied to the multipliers 150 and 151 on the input side of the mixers 121 and 120, respectively, to correct the offsets introduced by these mixers.

5   Although there are already many kinds of compensation circuits and methods, it is still a goal of research to propose newer and better solutions to the I/Q mismatch problem in transmitter systems.

SUMMARY OF THE INVENTION

10   The object of the present invention is to provide a new method and apparatus of I/Q channel mismatch and local leakage calibration in a transmitter.

The present invention provides a method for I/Q mismatch calibration of a transmitter. First, a discrete-time signal is
15   generated. Next, a corrected signal based on the discrete-time signal and a set of correction parameters A_p and B_p is obtained. Next, the corrected signal is converted to an analog corrected signal. Next, I/Q modulation is applied to the analog corrected signal and outputs a modulated signal. Next, a first desired
20   component measure and a first image component measure with a first set of the correction parameters A_p and B_p are obtained from the modulated signal. Next, a second desired component measure and a second image component measure with a second set of the correction parameters A_p and B_p are obtained from the modulated
25   signal. Next, a third desired component measure and a third image component measure with a third set of the correction parameters A_p and B_p are obtained from the modulated signal. Next, a fourth and fifth set of correction parameters A_p and B_p are obtained based on the first, the second, and the third desired component measures

as well as the first, the second, and the third image component measures. Next, a fourth desired component measure and a fourth image component measure with the fourth set of correction parameters A_p and B_p and a fifth desired component measure and a

5   fifth image component measure with the fifth set of correction parameters A_p and B_p are obtained from the modulated signal. Finally, a final set of the correction parameters A_p and B_p are selected from the fourth and fifth sets of correction parameters.

In addition, the method for I/Q mismatch calibration of a

10   transmitter according to the present invention further includes the following steps. A DC compensation parameter γ_p is added while obtaining the corrected signal $x_c[n]$ such that $x_c[n] = A_p \cdot (x[n] + \gamma_p) + B_p \cdot (x[n] + \gamma_p)^*$. Next, a first local leakage component measure is obtained from the modulated signal with the

15   final set of parameters A_p and B_p, and the first parameter $\gamma_p = \zeta_1$. A second local leakage component measure is then obtained from the modulated signal with the final set of parameters A_p and B_p, and the second parameter $\gamma_p = \zeta_2$. Next, a third local leakage component measure is obtained from the modulated signal with the

20   final set of parameters A_p and B_p, and the third parameter $\gamma_p = j\zeta_1$. A fourth local leakage component measure is then obtaining from the modulated signal with the final set of parameters A_p and B_p, and the fourth parameter $\gamma_p = j\zeta_2$. Next, a fifth local leakage component measure is obtaining from the modulated signal with the

25   final set of parameters A_p and B_p, and the fifth parameter $\gamma_p = 0$. Finally, a final DC compensation parameter is obtained based on the first local leakage component measure, the second local leakage component measure, the third local leakage component measure, the fourth local leakage component measure and the fifth

30   local leakage component measure.

In addition, the present invention further provides an apparatus for I/Q mismatch calibration of a transmitter. A signal generator generates a discrete-time signal. A correction module receives the discrete-time signal and obtains a corrected signal

5   based on the test signal and a set of correction parameters A_p and B_p. A first and second D/A converter converts the corrected signal to an analog signal, wherein the first D/A converter converts the real part of the corrected signal to the real part of the analog signal, and the second D/A converter converts the

10   imaginary part of the corrected signal to the imaginary part of the analog signal. A modulator applies I/Q modulation to the analog signal, and outputs a modulated signal. A measurer implements the steps of obtaining a first desired component measure and a first image component measure from the modulated

15   signal with a first set of the correction parameters A_p and B_p, obtaining a second desired component measure and a second image component measure from the modulated signal with a second set of the correction parameters A_p and B_p, obtaining a third desired component measure and a third image component measure from the

20   modulated signal with a third set of the correction parameters A_p and B_p, obtaining a fourth desired component measure and a fourth image component measure from the modulated signal with a fourth set of correction parameters A_p and B_p and obtaining a fifth desired component measure and a fifth image component measure

25   from the modulated signal with a fifth set of correction parameters A_p and B_p. A processor implements the steps of obtaining the fourth and fifth sets of correction parameters A_p and B_p based on the first, the second, and the third desired component measures as well as the first, the second, and the third

30   image component measures, and selecting a final set of correction

11

parameters A_p and B_p from the fourth and fifth sets of correction parameters.

BRIEF DESCRIPTION OF THE DRAWINGS

5   The present invention will become more fully understood from the detailed description given hereinbelow and the accompanying drawings, given by way of illustration only and thus not intended to be limitative of the present invention.

    FIG. 1 shows a quadrature modulator controlled by an

10   imbalance estimator as disclosed in U.S. Patent No. 6,208,698.

    FIG. 2 shows a setup of TX I/Q mismatch and carrier local leakage calibration for 802.11g of a transmitter.

    FIGs. 3A-3C are flowcharts of a method for I/Q mismatch and local leakage calibration of a receiver according to one

15   embodiment of the invention.

DETAILED DESCRIPTION OF THE INVENTION

    In the I/Q channel transmission architecture, there are two major sources of I/Q mismatch. One is the I/Q modulation and the

20   other is the I/Q transmission band filtering. The mismatch resulting from the I/Q modulation is modeled as a constant over the useful signal bandwidth, which is called systematic I/Q mismatch while the I/Q transmission band filtering causes frequency-dependent I/Q mismatch. The present invention

25   concentrates on the systematic I/Q mismatch. The mathematical theories of the invention will be described in the following.

    The baseband signals for the I and Q channels before the I/Q modulation are represented as $x_I(t)$ and $x_Q(t)$ respectively.

12

Assuming that an amplitude mismatch α, a phase mismatch θ, and zero local leakage results from the I/Q modulation, the equivalent baseband signals $y_I(t)$ and $y_Q(t)$ for the I and Q channels after the I/Q modulation are, without loss of
5   generality, given by the equations:

$$y_I(t) = x_I(t) \cdot (1+\alpha) \cdot \cos(\theta/2) + x_Q(t) \cdot (1-\alpha) \cdot \sin(\theta/2)$$
$$y_Q(t) = x_I(t) \cdot (1+\alpha) \cdot \sin(\theta/2) + x_Q(t) \cdot (1-\alpha) \cdot \cos(\theta/2) \qquad (7)$$

The combined baseband signal $x(t)$ before the I/Q modulation can be expressed as $x_I(t)+jx_Q(t)$ while the combined equivalent baseband signal $y(t)$ after the I/Q modulation can be expressed
10   as $y_I(t)+jy_Q(t)$. Thus, from the equation (7), the I/Q mismatch model can be expressed by the following equation:

$$y(t) = A \cdot x(t) + B \cdot x^*(t) \qquad (8)$$

where A and B are complex numbers with

$$A = \frac{1}{2}\left\{(1+\alpha)e^{j\theta/2} + (1-\alpha)e^{-j\theta/2}\right\}$$
$$B = \frac{1}{2}\left\{(1+\alpha)e^{j\theta/2} - (1-\alpha)e^{-j\theta/2}\right\} \qquad (9)$$

15   Eq. (8) is for the case of zero local leakage. When the local leakage γ is further considered, the I/Q mismatch model becomes

$$y(t) = A \cdot x(t) + B \cdot x^*(t) + \gamma \qquad (10)$$

One of solutions to the I/Q mismatch compensation is pre-distortion of the baseband signal $x(t)$ before the I/Q
20   modulation. If the signal is pre-distorted properly, the signal after the I/Q modulation will be free from the I/Q mismatch since the I/Q mismatch and the pre-distortion cancel out. Assuming that the baseband signal $x(t)$ is pre-distorted by a function D, the pre-distorted signal $z(t)$ is expressed as $D(x(t))$. The ideally
25   calibrated output signal should be:

$$y(t) = A \cdot D(x(t)) + B \cdot \left(D(x(t))\right)^* + \gamma \qquad (11)$$

$$= C \cdot x(t) \dotfill (12)$$

where C is a constant.

Accordingly, the I/Q mismatch compensation can be achieved by identifying the function D to satisfy the equation (12). One of solutions for the function D is a function with parameters A_p and B_p and is given by:

$$D(x(t)) = A_p \cdot x(t) + B_p \cdot x^*(t) \dotfill (13)$$

As will be shown, Eq.(13) will be implemented via digital signal process. The above is for the case of zero compenstation of local leakage. When considering the compensation of the local leakage of the IQ modulation, the function D becomes

$$D(x(t)) = A_p \cdot \left(x(t)+\gamma_p\right) + B_p \cdot \left(x(t)+\gamma_p\right)^* \dotfill (14)$$

By substituting the equation (14) into the equation (11), we derive:

$$y(t) = (A \cdot A_p + B \cdot B_p^*) \cdot x(t) + (A \cdot B_p + B \cdot A_p^*) \cdot x^*(t) \\ + (A \cdot A_p + B \cdot B_p^*) \cdot \gamma_p + (A \cdot B_p + B \cdot A_p^*) \cdot \gamma_p^* + \gamma \dotfill (15)$$

The equation (12) is satisfied if

$$\begin{cases} (A \cdot B_p + B \cdot A_p^*) = 0 \\ \gamma_p = -\gamma \end{cases} \dotfill (16)$$

Therefore, the goal of the calibration method in the invention is to identify the coefficients A_p, B_p, and γ_p in the equation (16).

FIG. 2 shows a setup of TX I/Q channel mismatch and carrier local leakage calibration for 802.11g of a transmitter according to the embodiment of the present invention. First, a discrete-time signal $x[n] = x(nT_s) = e^{j2\pi f_r nT_s}$ is generated by a signal generator 30, where $x(t) = e^{j2\pi f_r t}$ represents a single tone signal,

14

f_T is a real number representing the frequency in Hz of the signal $x(t)$ and T_s is the sampling period. Next, the signal $x[n]$ is passed to an I/Q correction module 31 including local leakage correction module 32 and gain/phase correction module 33, then the corrected

5 signal $x_c[n]$ output by the I/Q correction module 31 is fed to a pair of D/A converters 34, which convert the corrected signal $x_c[n]$ to an analog signal $x_c(t)$. To speak more specifically, the first D/A converter converts the real part of the corrected signal $x_c[n]$ to the real part of the analog signal $x_c(t)$ and the second

10 D/A converter converts the imaginary part of the corrected signal $x_c[n]$ to the imaginary part of the analog signal $x_c(t)$. The analog signal $x_c(t)$ is applied to an I/Q modulator, which increase the central frequency of the analog signal $x_c(t)$ by f_c Hz and outputting a modulated signal $x_m(t)$, wherein f_c is a preset real

15 number. The modulated signal $x_m(t)$ is then monitored by a spectrum analyzer 36, or other equipement that can monitor the signal specturm, to obtain the intensities of the frequency components of the modulated signal $x_m(t)$ at f_c+f_T, f_c-f_T, and f_c Hz, denoted by $W(f_T)$, $W(-f_T)$, and L, respectively. The $W(f_T)$, $W(-f_T)$, and L

20 are actually values indicative of the power of the equivalent baseband signal of $x_m(t)$ at f_T, $-f_T$, and 0 Hz, respectively. From the point of view of the I/Q correction, the term $W(f_T)$ represents a desired component, which will contain all the energy of the analog corrected signal $x_c(t)$ if the I/Q mismatch is ideally

25 compensated. And the term $W(-f_T)$ represents the image component, which is the additional interfering component at the image frequency due to the I/Q imbalance. And the term L represents the local leakage component, which is the baseband-equivalent DC component incurred by the carrier leakage. According to the

30 obtained data $W(f_T)$, $W(-f_T)$, and L, the software stored in a

15

personal computer (PC) 38 computes the parameters required by the I/Q correction module 31 to minimize the impacts due to the I/Q channel mismatch and local leakage. The software stored in the personal computer performs a different process from the prior

5 art.

FIGs. 3A-3C are flowcharts of a method for I/Q channel mismatch and local leakage calibration of a transmitter according to one embodiment of the invention.

First, a discrete-time signal $x[n] = x(nT_s) = e^{j2\pi f_T n T_s}$ is

10 generated by the signal generator 30 (S1), wherein $x(t) = e^{j2\pi f_T t}$ represents a single tone signal, f_T is a real number representing the frequency of the signal $x(t)$ and T_s is the sampling period. Next, in the present invention, the I/Q gain and phase errors are first calibrated. After the I/Q gain and phase errors are

15 corrected, the local leakage calibration procedure is then performed.

Let $x[n] = x_R[n] + jx_I[n]$ be the input signal of the I/Q correction module 31, which outputs a corrected signal $x_c[n]$ based on the discrete-time signal $x[n]$ and a set of correction parameters $(A_p,$

20 $B_p, \gamma_p)$ (S2). To speak more specifically, the output of the I/Q correction module 31 is expressed by $x_c[n] = A_P \cdot (x[n] + \gamma_P) + B_P \cdot (x[n] + \gamma_P)^*$

First, the corrected signal $x_c[n]$ output by the I/Q correction module 31 is fed to a pair of D/A converters 34, which then convert the corrected signal $x_c[n]$ to an analog signal $x_c(t)$.

25 To speak more specifically, the first D/A converter converts the real part of the corrected signal $x_c[n]$ to the real part of the analog signal $x_c(t)$ and the second D/A converter converts the imaginary part of the corrected signal $x_c[n]$ to the imaginary part of the analog signal $x_c(t)$. The analog signal $x_c(t)$ is then applied

30 to an I/Q modulator, which increases the central frequency of the

analog signal $x_c(t)$ by f_c Hz and outputs a modulated signal $x_m(t)$, wherein f_c is a preset real number. The modulated signal $x_m(t)$ is then analyzed by the spectrum analyzer 36. In the present invention, a plurality of sets of parameters A_p, B_p and γ_p are

5   predetermined, and then the spectrum analyzer 36 analyzes the modulated signal $x_m(t)$ with different sets of the parameters A_p, B_p and γ_p.

In the I/Q mismatch calibration procedure (S3~S8) according to the embodiment of the present invention, the parameter γ_p is

10   zero during the procedure. A first desired component measure $W^{(1)}(f_T)$ and a first image component measure $W^{(1)}(-f_T)$ is obtained with a first set of the correction parameters A_p and B_p (S3). The first set of the correction parameters A_p and B_p is $(A_p, B_p) = (a, 0)$, wherein a is a real number.

15   When $a=1$, we have $x_c(t) = x(t)$. The equivalent baseband signal of the modulated signal, denoted by $y(t)$, is given by :

$$\begin{aligned} y(t) &= A \cdot x[n] + B \cdot x^*[n] + \gamma \\ &= A \cdot e^{j2\pi f_T t} + B \cdot e^{-j2\pi f_T t} + \gamma \end{aligned} \quad \text{.......................................(17)}$$

thus, the intensities of the resulting desired tone component $W^{(1)}(f_T)$, the image component $W^{(1)}(-f_T)$ associated with

20   the first set of the correction parameters A_p and B_p are obtained from the spectrum analyzer 36, see the equation (18).

$$W^{(1)}(f_T) = \beta_1 \, |A|^2 = \beta \left| \cos\frac{\theta}{2} + j\alpha\sin\frac{\theta}{2} \right|^2 = \beta \left(\cos^2\frac{\theta}{2} + \alpha^2 \sin^2\frac{\theta}{2} \right)$$
$$\text{...............(18)}$$
$$W^{(1)}(-f_T) = \beta_1 \, |B|^2 = \beta \left| \alpha\cos\frac{\theta}{2} + j\sin\frac{\theta}{2} \right|^2 = \beta \left(\alpha^2 \cos^2\frac{\theta}{2} + \sin^2\frac{\theta}{2} \right)$$

where β_1 is an unknown constant related to the spectrum analyzer

25   36. Here, the measures $W^{(1)}(f_T)$ and $W^{(1)}(-f_T)$ are values indicative of the power of the signal $y(t)$ at f_T and $-f_T$ Hz, respectively.

17

In this case, the image-rejection ratio $\rho^{(1)}$ is

$$\rho^{(1)} = \frac{W^{(1)}(-f_T)}{W^{(1)}(f_T)} = \frac{\alpha^2 \cos^2\frac{\theta}{2} + \sin^2\frac{\theta}{2}}{\cos^2\frac{\theta}{2} + \alpha^2 \sin^2\frac{\theta}{2}} \quad\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (19)$$

Next, a second desired component measure $W^{(2)}(f_T)$ and a second image component measure $W^{(2)}(-f_T)$ are obtained from the spectrum analyzer 36 with a second set of the correction parameters A_p and B_p (S4). The second set of the correction parameters A_p and B_p is $(A_p, B_p) = (b, b)$, wherein 'b' is a real number.

When $b = \dfrac{1}{2}$:

$$\begin{aligned} y(t) &= A \cdot \left(0.5 \cdot x(t) + 0.5 \cdot x^*(t)\right) + B \cdot \left(0.5 \cdot x(t) + 0.5 \cdot x^*(t)\right)^* + \gamma \\ &= \left(0.5A + 0.5B\right) \cdot x(t) + \left(0.5A + 0.5B\right) \cdot x^*(t) + \gamma \qquad\ldots\ldots\ldots\ldots\ldots\ldots (20) \\ &= \left(0.5A + 0.5B\right) \cdot e^{j2\pi f_T t} + \left(0.5A + 0.5B\right) \cdot e^{-j2\pi f_T t} + \gamma \end{aligned}$$

Thus, the intensities of the resulting desired tone component $W^{(2)}(f_T)$, the image component $W^{(2)}(-f_T)$ are obtained from the spectrum analyzer 36 with the second set of the correction parameters A_p and B_p , see the equation (21).

$$\begin{aligned} W^{(2)}(f_T) &= \beta_2 \mid 0.5A + 0.5B \mid^2 = \frac{\beta_2}{4}\left|\cos\frac{\theta}{2} + j\alpha\sin\frac{\theta}{2} + \alpha\cos\frac{\theta}{2} + j\sin\frac{\theta}{2}\right|^2 \\ &= \frac{\beta_2}{4}\left|(1+\alpha)\cos\frac{\theta}{2} + j(1+\alpha)\sin\frac{\theta}{2}\right|^2 \\ &= \frac{\beta_2}{4}\left((1+\alpha)^2\cos^2\frac{\theta}{2} + (1+\alpha)^2\sin^2\frac{\theta}{2}\right) \\ &= \frac{\beta_2(1+\alpha)^2}{4} \end{aligned}$$

$$W^{(2)}(-f_T) = \beta_2 \mid 0.5A + 0.5B \mid^2 = \frac{\beta_2(1+\alpha)^2}{4}$$

(21)

Here, from the output of the spectrum analyzer 36, a value N= $W^{(2)}(-f_T) = W^{(2)}(f_T)$ is proportional to $(1+\alpha)^2$.

Next, a third desired component measure $W^{(3)}(f_T)$ and a third image component measure $W^{(3)}(-f_T)$ are obtained from the spectrum analyzer 36 with a third set of the correction parameters A_p and B_p (S5). The third set of the correction parameters A_p and B_p is $(A_p, B_p) = (b, -b)$.

When $b=\dfrac{1}{2}$:

$$y(t) = A\left(0.5\cdot x(t) - 0.5\cdot x^*(t)\right) + B\left(0.5\cdot x(t) - 0.5\cdot x^*(t)\right)^* + \gamma$$
$$= \left(0.5A - 0.5B\right)\cdot x(t) + \left(-0.5A + 0.5B\right)\cdot x^*(t) + \gamma \qquad\qquad \text{............ (22)}$$
$$= \left(0.5A - 0.5B\right)\cdot e^{j2\pi f_T t} + \left(-0.5A + 0.5B\right)\cdot e^{-j2\pi f_T t} + \gamma$$

Thus, the intensities of the resulting desired tone component $W^{(3)}(f_T)$, the image component $W^{(3)}(-f_T)$ are obtained from the spectrum analyzer 36 with the third set of the correction parameters A_p and B_p, see the equation (23).

$$W^{(3)}(f_T) = \beta_2 \mid 0.5A - 0.5B \mid^2 = \frac{\beta_2}{4}\left|\cos\frac{\theta}{2} + j\alpha\sin\frac{\theta}{2} - \alpha\cos\frac{\theta}{2} - j\sin\frac{\theta}{2}\right|^2$$
$$= \frac{\beta_2}{4}\left|(1-\alpha)\cos\frac{\theta}{2} + j(\alpha-1)\sin\frac{\theta}{2}\right|^2$$
$$= \frac{\beta_2}{4}\left((1-\alpha)^2 \cos^2\frac{\theta}{2} + (1-\alpha)^2 \sin^2\frac{\theta}{2}\right) \qquad\qquad (23)$$
$$= \frac{\beta_2(1-\alpha)^2}{4}$$

$$W^{(3)}(-f_T) = \beta_2 \mid -0.5A + 0.5B \mid^2 = \frac{\beta_2(1-\alpha)^2}{4}$$

where β_2 is an unknown constant related to the spectrum analyzer 36. Here, from the output of the spectrum analyzer 36, a value

O= $W^{(3)}(-f_T)=W^{(3)}(f_T)$ is proportional to $(1-\alpha)^2$. Thus, the gain mismatch term α can be estimated by

$$\alpha \approx \hat{\alpha} = \frac{\sqrt{N/O}-1}{\sqrt{N/O}+1} \quad\text{..}(24)$$

Term N can be set to be either $N=W^{(2)}(f_T)$ or $N=W^{(2)}(-f_T)$. Due to the existence of the measurement error, the term N can also

set to be $N=(W^{(2)}(f_T)+W^{(2)}(-f_T))/2$ or $N=\sqrt{W^{(2)}(f_T)\cdot W^{(2)}(-f_T)}$ to

increase the accuracy. Similarly, Term O can be set to be either $O=W^{(3)}(f_T)$ or $O=W^{(3)}(-f_T)$. Due to the existence of the measurement error, the term O can also set to be $O=(W^{(3)}(f_T)+W^{(3)}(-f_T))/2$ or

$O=\sqrt{W^{(3)}(f_T)\cdot W^{(3)}(-f_T)}$ to increase the accuracy. Substitute α into

the equation (19), the values Q and P are obtained, see the equation (25).

$$Q = \frac{\hat{\alpha}^2-\rho^{(1)}}{\left(1+\rho^{(1)}\right)(\hat{\alpha}^2-1)} = \sin^2\frac{\theta}{2} \quad\text{...}(25)$$

$$P = 1-Q = 1-\sin^2\frac{\theta}{2} = \cos^2\frac{\theta}{2} \quad\text{...}(26)$$

Next, a fourth and fifth set of correction parameters A_p and B_p are obtained based on the first desired component measure $W^{(1)}(f_T)$, the second desired component measure $W^{(2)}(f_T)$, and the third desired component measure $W^{(3)}(f_T)$, and the first image component measure $W^{(1)}(-f_T)$, the second image component measure $W^{(2)}(-f_T)$, and the third image component measure $W^{(3)}(-f_T)$ (S6).

The fourth and fifth sets of correction parameters A_p and B_p are expressed by parameters $\hat{\alpha}$, P, and Q. The fourth set of the correction parameters A_p and B_p is

20

$$A_p = \sqrt{P} - j\hat{\alpha}\sqrt{Q}$$
$$B_p = -\hat{\alpha}\sqrt{P} - j\sqrt{Q} \quad ,$$

and the fifth set of the correction parameters A_p and B_p is

$$A_p = \sqrt{P} + j\hat{\alpha}\sqrt{Q}$$
$$B_p = -\hat{\alpha}\sqrt{P} + j\sqrt{Q} \quad .$$

Next, a fourth desired component measure $W^{(4)}(f_T)$ and a

5   fourth image component measure $W^{(4)}(-f_T)$ are obtained with the
fourth set of correction parameters A_p and B_p, and a fifth desired
component measure $W^{(5)}(f_T)$ and a fifth image component measure
$W^{(5)}(-f_T)$ are obtained with the fifth set of correction parameters
A_p and B_p (S7).

10   Last, a final set of the correction parameters A_p and B_p is
determined according to the fourth and fifth sets of correction
parameters (S8). In addition, since there is no information about
the sign of the phase mismatch θ, a hypothesis testing is

performed to determine whether $\sin\dfrac{\theta}{2} = \sqrt{Q}$ or $\sin\dfrac{\theta}{2} = -\sqrt{Q}$, however,

15   in either case, $\cos\dfrac{\theta}{2} = \sqrt{1-Q}$ since $-\dfrac{\pi}{2} < \theta < \dfrac{\pi}{2}$ in typical

conditions.

If the image rejection ratio corresponding to the fourth set
of (A_p, B_p) is less than or equal to the image rejection ratio
corresponding to the fifth set of (A_p, B_p), the fourth set of (A_p, B_p)

20   is chosen as the final set of (A_p, B_p), otherwise, the fifth set
of (A_p, B_p) is chosen as the final set of (A_p, B_p). In addition,
the final set of (A_p, B_p) can be further multiplied by a

normalization factor $\chi = \dfrac{1}{AA_p + BB_p^*} = \dfrac{1}{(1-\hat{\alpha}^2)\cdot(1-2Q)}$ to ensure the

unity gain of the correction process. The steps S6 and S8 are performed by the personal computer 38 shown in FIG. 3 of the present invention.

In addition, it is not the only way to determine the final set of the correction parameters A_p and B_p through the value of $\dfrac{W^{(4)}(-f_T)}{W^{(4)}(f_T)}$ and $\dfrac{W^{(5)}(-f_T)}{W^{(5)}(f_T)}$. Under the final parameter, the image component would be very small (theoretically it is zero), and therefore would be much less than the image component under the other sets of parameter. From another angle, a small image component will imply a strong desired component, since their sum should be the same in split regardless of the set of parameters. Therefore, under the final parameter, the desired component would be stronger than the desired component under the other set of parameters.

Next, the local leakage calibration is active after the elimination of I/Q channel mismatch errors in step S9~S13.

In the local leakage calibration procedure according to the embodiment of the present invention, a first local leakage component measure L_1 of the corrected signal is obtained from the spectrum analyzer 36 with the final set of parameters A_p and B_p, and the parameter $\gamma_p = \zeta_1$ (S9), where ζ_1 is a real number.

Thus, the carrier leakage term at the spectrum analyzer 36 is

$$L_1 = \beta_3 \cdot (\gamma_R^2 + \gamma_I^2 + \zeta_1^2 + 2\gamma_R \cdot \zeta_1) \dotfill (27)$$

where β_3 is a scaling factor.

Next, a second local leakage component measure L_2 of the corrected signal is obtained from the spectrum analyzer 36 with the final set of parameters A_p and B_p, and the parameter $\gamma_p = \zeta_2$ (S10), where ζ_2 is a real number.

Thus, the carrier leakage term at the spectrum analyzer 36 is

$$L_2 = \beta_3 \cdot (\gamma_R^2 + \gamma_I^2 + \zeta_2^2 + 2\gamma_R \cdot \zeta_2) \ . \quad\text{..(28)}$$

Next, a third local leakage component measure L_3 of the corrected signal is obtained from the spectrum analyzer 36 with the final set of parameters A_p and B_p, and the parameter $\gamma_p = j\zeta_1$ (S11), where ζ_1 is a real number.

Thus, the carrier leakage term at the spectrum analyzer 36 is

$$L_3 = \beta_3 \cdot (\gamma_R^2 + \gamma_I^2 + \zeta_1^2 + 2\gamma_I \cdot \zeta_1) \ . \quad\text{..(29)}$$

Next, a fourth local leakage component measure L_4 of the corrected signal is obtained from the spectrum analyzer 36 with the final set of parameters A_p and B_p, and the parameter $\gamma_p = j\zeta_2$ (S12), where ζ_2 is another real number.

Thus, the carrier leakage term at the spectrum analyzer 36 is

$$L_4 = \beta_3 \cdot (\gamma_R^2 + \gamma_I^2 + \zeta_2^2 + 2\gamma_I \cdot \zeta_2) \ . \quad\text{..(30)}$$

Next, a fifth local leakage component measure L_5 of the corrected signal is obtained from the spectrum analyzer 36 with the final set of parameters A_p and B_p, and the parameter $\gamma_p = 0$ (S13).

Thus, the carrier leakage term at the spectrum analyzer 36 is

$$L_5 = \beta_3 \cdot (\gamma_R^2 + \gamma_I^2) \ . \quad\text{..(31)}$$

Thus, the parameters γ_R and γ_I are obtained based on the first local leakage component measure L_1, the second local leakage component measure L_2, the third local leakage component measure L_3, the fourth local leakage component measure L_4 and the fifth

local leakage component measure L_5 (S14), see the equations (32) and (33).

$$\hat{\gamma}_R = -\frac{1}{2}\frac{\zeta_2^2(L_1-L_5)-\zeta_1^2(L_2-L_5)}{\zeta_1(L_2-L_5)-\zeta_2(L_1-L_5)} \quad\text{..}(32)$$

$$\hat{\gamma}_I = -\frac{1}{2}\frac{\zeta_2^2(L_3-L_5)-\zeta_1^2(L_4-L_5)}{\zeta_1(L_4-L_5)-\zeta_2(L_3-L_5)}\text{..}(33)$$

Finally, according to the step S14, the final DC compensation parameter $\gamma_{p,final}$ is obtained by

$$\gamma_{p,final} = -\frac{1}{2}\cdot\frac{\zeta_2^2(L_1-L_5)-\zeta_1^2(L_2-L_5)}{\zeta_1(L_2-L_5)-\zeta_2(L_1-L_5)} - j\frac{1}{2}\cdot\frac{\zeta_2^2(L_3-L_5)-\zeta_1^2(L_4-L_5)}{\zeta_1(L_4-L_5)-\zeta_2(L_3-L_5)}\text{............}(34)$$

In addition, the step S14 is performed by the personal computer 38 shown in FIG. 2 of the present invention.

The previously described method can be applied to a transceiver module, e.g., an IEEE 802.11 compliant wireless LAN transceiver module, involving I/Q demodulation. Wireless LANs based on the IEEE 802.11 standard have achieved wide customer acceptance in the enterprise environment. They are expected to continue to expand in popularity and become ubiquitous communication systems even in private and public places.

Thus, after the calibration procedure, the I/Q correction module compensates the I/Q channel mismatch in the signal from the baseband transmitter by the finally determined coefficients A_p, B_p and γ_p. The previously described calibration procedure may be implemented upon production stage of the transceiver.

In conclusion, the present invention provides a new method and apparatus for transmitter I/Q mismatch calibration, especially suitable for an IEEE 802.11g compliant WLAN transceiver module. The compensation of the I/Q mismatch is achieved by a signal pre-distortion prior to the I/Q modulation. The coefficients of the pre-distortion function are determined

by a calibration procedure. After performing the procedure disclosed by the embodiment of the present invention, the coefficients are accurate enough for the transmitted signal to be free from the I/Q mismatch.

5   In addition, the present invention can be applied to any kind of transmitter having an I/Q modulator, and the OFDM transmitter is just one example, thus it is not limited to the scope of the present invention.

The foregoing description of the preferred embodiments of

10   this invention has been presented for purposes of illustration and description. Obvious modifications or variations are possible in light of the above teaching. The embodiments were chosen and described to provide the best illustration of the principles of this invention and its practical application to

15   thereby enable those skilled in the art to utilize the invention in various embodiments and with various modifications as are suited to the particular use contemplated. All such modifications and variations are within the scope of the present invention as determined by the appended claims when interpreted

20   in accordance with the breadth to which they are fairly, legally, and equitably entitled.